Introduction to RADAR systems

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Merrill I. Skolnik

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chapter

11

Radar Receiver

11.1 THE RADAR RECEIVER

The function of the receiver in early radar systems was to extract the weak echo signals that appeared at the antenna terminals and amplify them to a level where they could be displayed to a radar operator who then made the decision as to whether or not a target echo signal was present. The modern radar receiver still has to extract the weak echo signals and amplify them, but it does much more. It employs a matched filter (Sec. 5.2) whose purpose is to maximize the peak-signal-to-mean-noise-ratio and discriminate against unwanted signals whose waveforms are different from those transmitted by the radar. When the clutter echoes are large enough to mask desired target echoes, the receiver also has to incorporate means for separating the moving targets from stationary clutter echoes by recognizing the doppler frequency shift of the moving targets (Chap. 3).

In modern radars the decision whether a target is present or absent is seldom made by an operator viewing on a display the unprocessed output of a receiver. Instead, the detection decision is made automatically based on threshold detection (Sec. 5.5). Information about a target's location in range and angle can also be extracted automatically instead of manually by an operator. In an operational air-surveillance radar, tracking of targets is no longer performed by an operator marking with a grease pencil on a radar display the location of blips (targets) from scan to scan and calculating the target speed and

estimating its direction. Targets are acquired and tracked automatically (Sec. 4.9) and only processed tracks are displayed to the operator or sent to some automatic device, such as an air-traffic control system or weapons control computer, for further use. When the radar cannot remove all the clutter echoes, constant false alarm rate (CFAR) circuitry is employed to prevent the tracking computer from becoming overloaded when trying to establish tracks using clutter echoes. The receiver is also the place where external interference and hostile electronic-countermeasures are kept from interfering with the detection of targets.

Thus, in addition to detection and amplification of signals, a radar receiver performs many other functions either directly as part of the receiver or in conjunction with it. These other functions include signal processing, information extraction, data processing, electromagnetic compatibility, and electronic counter-countermeasures. (The modern receiver might be thought of as the *receiver/processor.*) Sometimes the display is considered part of the receiver system. In this and other radar books, these other functions are often considered separately from the discussion of the receiver. The interested reader will find a more thorough review of the radar receiver by John W. Taylor, Jr. in the *Radar Handbook.*[1]

The radar receiver is almost always a *superheterodyne,* or superhet. It was shown in the block diagram of Fig. 1.4 and briefly described in Sec. 1.3. The essential characteristic of a superheterodyne is that it converts the RF input signal to an intermediate frequency (IF) where it is easier than at RF to achieve the necessary filter shape, bandwidth, gain, and stability. An advantage of the superheterodyne receiver is that its frequency can be readily changed by changing the frequency of the local oscillator (LO). The first stage, or *front-end,* of a radar superheterodyne receiver can be an RF low-noise amplifier (LNA) such as a transistor.

Before the availability of low-noise transistors, the receiver front-end was the mixer stage without an RF amplifier preceding it. In some applications the mixer stage might still be desired as the receiver front-end instead of a low-noise amplifier. A receiver with a mixer as the first stage has a greater dynamic range than one with a low-noise amplifier, which might be important when large MTI improvement factors are needed to remove clutter echoes. The extra dynamic range available with a mixer as the front-end can also be of value in reducing the likelihood of receiver saturation when large signals or jamming are present. The larger receiver noise figure of a mixer might be compensated with greater transmitter power and/or a bigger antenna, both of which are beneficial when a military radar is faced with hostile noise jamming. Although the mixer front-end might have some advantage over a low-noise transistor amplifier front-end, a low-noise amplifier as the first stage of a superheterodyne receiver generally seems to be preferred by those who buy radars.

A high-performance air-surveillance radar sometimes employs more than one type of receiver. Each would share the front-end, mixer, and IF stages. One receiver might be a linear amplifier and envelope detector for detection of targets in the clear (no competing clutter). A second receiver might be for doppler processing to remove clutter, as in an MTI radar. It would use I and Q channels and digital signal processing to filter the moving targets. A third receiver might be a log-FTC (Sec. 7.8), or something similar, to aid in detecting targets located within moving weather clutter beyond the range of surface clutter.

It was said in Chap. 2 that if the radar designer wishes to increase the detection range of a radar, the chief factors available are the average power of the transmitter and the area of the antenna. The classical radar equation also indicates that the range can be increased by reducing the receiver noise figure; but, in practice, the noise figures of radar receivers are already quite low and any further decrease can produce marginal results and, sometimes, unwanted effects. Further lowering of the noise figure might not be justified if it significantly increases receiver cost, lowers the dynamic range, subjects the device to a greater risk of burn-out, and results in less reliability. A very sensitive receiver also allows more interference to enter the receiver. Sometimes an increase in interference is a price that may be worth the benefits of improved sensitivity, but there can be limits.

A radar receiver has to have sufficient gain to increase the level of the weak echo signal to where it is large enough to be processed or displayed. In the superheterodyne the total receiver gain is divided between the IF and the video amplifiers. The receiver should have adequate dynamic range (where the receiver is linear) so that large clutter echoes do not cause the receiver to saturate and reduce the MTI improvement factor. It must not introduce unwanted phase or amplitude changes that could distort the echo signals. It must be protected from overload, saturation, and damage (burnout) by strong unwanted signals. Timing and reference signals are needed to properly extract target information and take advantage of the doppler frequency shift of echo signals.

A limitation with early radar receivers when vacuum tubes were the only available technology was that they were relatively large in size. The size of radar receivers is no longer a problem with modern technology. The trend is to make the receiver as digital as is practical, with analog components confined to the RF and perhaps the IF.

There can be many demands on the radar receiver, but the receiver designer has responded well to the challenges and there exists a highly developed state of receiver technology. Radar receiver design and implementation may not always be an easy task, but receiver designers have usually been able to provide the radar systems engineer the means to accomplish the desired objectives.

11.2 RECEIVER NOISE FIGURE[2,3]

Definition The receiver noise figure was described in Sec. 2.3 as a measure of the noise produced by a practical receiver compared to the noise of an ideal receiver. The noise figure of a linear network may be defined as either

$$F_n = \frac{N_{out}}{kT_0B_nG} \quad \text{or} \quad \frac{S_{in}/N_{in}}{S_{out}/N_{out}} \qquad [11.1]$$

where N_{out} = available output noise power; $kT_0B_n = N_{in}$ = available input noise power; k = Boltzmann's constant = 1.38×10^{-23} J/deg; T_0 = standard temperature of 290 K (approximately room temperature); B_n = noise bandwidth defined by Eq. (2.3); $G = S_{out}/S_{in}$ = available gain; S_{out} = available output signal power; and S_{in} = available input signal power. The term "available power" refers to the power that would be delivered to a matched load. (The term "available" will be understood in the following discussion of noise figure and is not mentioned further.) The product $kT_0 = 4 \times 10^{-21}$ W/Hz.

The reason for a standard temperature T_0 in the definition of noise figure is to refer measurements made under differing temperature conditions to a common basis of comparison.

Equation (11.1) permits two different, but equivalent, interpretations of the noise figure. It may be considered (right-hand side) as the degradation of the signal-to-noise ratio as the signal passes through the network, or (left-hand side) it may be interpreted as the ratio of the noise-power out of the actual network to the noise-power out of an ideal network that amplifies the input thermal noise and introduces no additional noise of its own. The noise figure of Eq. (11.1) can be expanded as

$$F_n = \frac{kT_0B_nG + \Delta N}{kT_0B_nG} = 1 + \frac{\Delta N}{kT_0B_nG} \qquad [11.2]$$

where ΔN is the additional noise introduced by the practical (nonideal) network.

The noise figure is commonly expressed in decibels; that is, 10 log F_n. The term *noise factor* has also been used at times instead of noise figure. The definition of noise figure assumes that the input and output of the network are matched. In some devices, less noise is obtained under mismatched, rather than matched, conditions. In spite of definitions, such networks would be operated so as to achieve the maximum output signal-to-noise ratio.

Noise Figure of Networks in Cascade Consider two networks in cascade, each with the same noise bandwidth B_n but with different noise figures and gain, Fig. 11.1. Let F_1, G_1 be the noise figure and gain, respectively, of the first network and F_2, G_2 be similar parameters for the second network. The problem is to find F_o, the overall noise-figure of the two networks in cascade. From the definition of noise figure given by Eqs. (11.1) and (11.2), the output noise N_{out} of the two networks in cascade is

$$N_{out} = \text{noise from network 1 at output of network 2} + \text{noise } \Delta N_2 \text{ introduced by network 2}$$

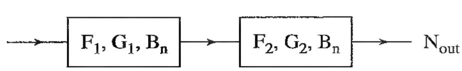
$$= F_0kT_0B_nG_1G_2 = F_1kT_0B_nG_1G_2 + \Delta N_2 = F_1kT_0B_nG_1G_2 + (F_2 - 1)\,kT_0B_nG_2$$

which results in

$$F_0 = F_1 + \frac{F_2 - 1}{G_1} \qquad [11.3]$$

It is not sufficient that the first stage of a low-noise receiver have a low noise figure. The second stage must also have a low noise figure or, if not, the gain of the first stage needs to be large. Too large a first-stage gain, however, is not always desirable since the dynamic range of the receiver is reduced by the gain G_1 of the low-noise amplifier. If the first network is not an amplifier, but is a diode mixer, the gain G_1 should be interpreted as a number less than unity (a loss).

Figure 11.1 Two networks in cascade with different noise figures and gains, but the same noise bandwidths.



The noise figure of N networks in cascade may be shown to be

$$F_0 = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \cdots + \frac{F_N - 1}{G_1 G_2 \cdots G_{N-1}} \qquad [11.4]$$

Similar expressions may be derived when the bandwidth and/or temperature of the individual networks are not the same.[4]

Noise Figure Due to Loss in the Transmission Line Any losses in the RF portion ahead of the receiver front-end result in an increase in the apparent overall noise figure. Such losses can be due to the transmission line between antenna and receiver, the duplexer, receiver protector, rotary joint, preselector filter, STC if applied at RF, monitoring devices, and the radome. The noise figure due to these RF losses, obtained from the second part of the definition of Eq. (11.1), is equal to the RF loss L_{RF}. (This can be seen since L_{RF} is the loss in signal-to-noise ratio as the signal travels from the antenna to the receiver.) It can also be obtained from the first part of Eq. (11.1) since the noise out of a lossy transmission line is kT_0B_n, and its gain $G = 1/L_{RF}$.

If the loss in the transmission line and its associated devices is incorporated in the receiver noise figure, it should not also be a part of the system losses. Most radar analyses treat these losses as system losses rather than as part of the receiver noise figure. There are some RF devices that are likely to be closely associated with the low-noise receiver. These include the circulator that provides isolation, the receiver protector, waveguide to coax transition, and receiver performance monitoring (which might produce a loss). When a noise figure for a receiver is quoted in a publication or given in a manufacturer's catalog, it might not always be obvious what is included. It might be the receiver noise figure without the associated receiver losses or it might include the loss of those devices that are a close part of the receiver, as mentioned above. With a mixer front-end receiver the noise figure quoted has sometimes been that of the mixer alone and not of the entire receiver. There seems to be no standard method for reporting receiver noise figure.

Noise Temperature The noise introduced by a network may also be expressed as the *effective noise temperature*, T_e, defined as the (fictional) temperature at the input of the network, that accounts for the additional noise ΔN at the output. Therefore $\Delta N = kT_eB_nG$ and from Eq. (11.2) we have

$$F_n = 1 + \frac{T_e}{T_0} \qquad [11.5]$$

$$T_e = (F_n - 1)T_0 \qquad [11.6]$$

The *system noise temperature* T_s is defined as the effective noise temperature of the receiver including the effects of antenna temperature T_a. If the receiver effective noise temperature is T_e, then

$$T_s = T_a + T_e = (F_s - 1)T_0 \qquad [11.7]$$

where F_s is the *system noise figure*. This equation also defines the system noise figure when it includes the effects of the antenna temperature T_a and receiver effective noise temperature T_e.

The effective noise temperature of a receiver consisting of a number of networks in cascade is

$$T_e = T_1 + \frac{T_2}{G_1} + \frac{T_3}{G_1 G_2} + \cdots \qquad [11.8]$$

where T_i and G_i are the effective noise temperature and gain of the ith network.

The effective noise temperature and the noise figure both describe the same characteristic of a network. The effective noise temperature generally is used to describe the noise performance of very low-noise receivers, lower than might be of interest for radar. It is also preferred by some radar engineers and many receiver designers as being more useful than noise figure for analysis purposes. The noise figure, however, seems to be the more widely used term to describe radar receiver performance, and is used in this text for that purpose.

11.3 SUPERHETERODYNE RECEIVER

The discussion of the superheterodyne receiver in this section does not include all aspects of the receiver, but only with those component parts that have an effect on the radar system design. This includes the low-noise RF amplifier, the mixer, receiver dynamic range, the $1/f$ noise at IF, oscillator noise, and the detector.

Low-Noise Front-End The first stage of a superheterodyne receiver for radar application can be a transistor amplifier. At the lower radar frequencies the silicon bipolar transistor has been used. Gallium-arsenide field-effect transistors (FET) are found at the higher frequencies. Other types of transistors also can be used, depending on the trade-off between the desired noise figure and the ability of the transistor to withstand burnout. An X-band transistor can provide a noise figure of about one dB and can withstand a leakage peak power of 0.2 W.[5] With a diode limiter ahead of the transistor, the peak power can be as great as 50 W before burnout. The diode limiter increases the noise figure about 0.5 dB at X band and 0.2 dB at C band. The lower the frequency the lower can be the transistor noise figure. At C band the noise figure might be around 0.6 dB. These values are more than adequate for radar. (Early microwave radars had noise figures of 12 to 15 dB and radars in the 1960s had noise figures of 7 to 8 dB.) It is not necessary for the radar systems engineer to have extremely low noise figures in most radar applications, especially when the unavoidable losses in the transmission line between receiver and antenna are considered. If improved radar system performance is of concern, it is probably more fruitful to try to reduce some of the many system losses that occur elsewhere in a radar rather than try to reduce further the noise figure of the low-noise amplifier (LNA). It is usually good enough.

Prior to the low-noise transistor amplifier, the parametric amplifier and the maser were available as low-noise receiver front-ends. Although their noise figures were low (lower than those of transistors, which came later), they were seldom used operationally for radar. They were expensive, of large size, and often did not have sufficient dynamic

range. Until low noise transistor amplifiers were developed, the radar receiver seldom employed an RF amplifier stage except perhaps at UHF or lower frequencies. Before the low-noise transistor, the mixer was the receiver front-end. As already mentioned, a mixer as the front-end without a low-noise amplifier preceding it is a valid option for some radar applications in spite of its higher receiver noise figure.

Achieving low receiver noise is no longer the problem it once was, and designers of high-performance radar receivers usually are more concerned with obtaining large dynamic range and low oscillator noise.

Mixers[6] Whether or not it is used as the front-end, the mixer is a key element in a superheterodyne receiver since it is the means by which the incoming RF signal is converted to IF (intermediate frequency). When the down conversion from RF to IF is performed in one step, it is called single conversion. Sometimes the down conversion is done in two steps with two mixers and IF amplifiers. This is known as dual conversion. Dual conversion superheterodyne receivers are used to avoid some forms of interference and (spoofing) electronic countermeasures. The mixer should have a low conversion loss, introduce little additional noise of its own, minimize spurious responses, and not be susceptible to burnout, especially when it is used as the front-end without a low noise amplifier ahead of it. An integral part of the mixer is the local oscillator.

Noise Figure of a Mixer Used as a Front End The noise figure of a mixer is determined by its conversion loss and noise-temperature ratio. The conversion loss of a mixer is defined as

$$L_c = \frac{\text{available RF power}}{\text{available IF power}} \tag{11.9}$$

It is a measure of the efficiency of the mixer in converting RF signal power into IF. The conversion loss of typical microwave diodes in a conventional single-ended mixer configuration varies from about 5 to 6.5 dB. Schottky diodes in an image recovery mixer have been reported to have a minimum conversion loss of 3.5 dB over a narrow band and a 4-dB conversion loss over a 10 percent bandwidth at S band.[7] The noise-temperature ratio of a mixer (not to be confused with the effective noise temperature) is defined as

$$t_r = \frac{\text{actual available IF noise power}}{\text{available noise power from an equivalent resistance}} \tag{11.10}$$

or

$$t_r = \frac{F_m k T_0 B G_c}{k T_0 B} = F_m G_c = \frac{F_m}{L_c}$$

where F_m = mixer noise figure and L_c = conversion loss = $1/G_c$. The noise temperature ratio of a mixer varies inversely with the IF frequency from about 100 kHz down to a small fraction of a hertz. At a frequency of 30 MHz, the noise temperature ratio might range from 1.2 to 2.0. Generally, the lower the conversion loss, the larger is the noise temperature ratio.

The noise figure of a mixer based on Eq. (11.10) is $F_m = L_c t_r$. It is, however, not a complete measure of the sensitivity of a receiver with a mixer front-end. The overall noise figure depends not only on the mixer stage, but also on the noise figure of the IF amplifier. The latter becomes a significant factor in the overall noise figure since the mixer has a loss rather than a gain. Using Eq. (11.3), the noise figure of the first network (the mixer) is $F_1 = L_c t_r$ and its gain is $G_1 = 1/L_c$. The noise figure of the second network is that of the IF amplifier, so that $F_2 = F_{\text{IF}}$. The receiver noise figure with a mixer front-end is then

$$F_R = F_1 + \frac{F_2 - 1}{G_1} = L_c t_r + (F_{\text{IF}} - 1)L_c = L_c(t_r + F_{\text{IF}} - 1) \qquad [11.11]$$

(This does not include any losses in the RF transmission line.) If, for example, the conversion loss of the mixer were 5.5 dB, the IF noise figure 0.5 dB, and the noise-temperature ratio 1.2, the receiver noise figure would be 6.7 dB. For low noise-temperature diodes, the receiver noise figure is approximately equal to the conversion loss times the IF noise figure.

Some manufacturers have used Eq. (11.11) to determine the noise figures of the mixers listed in their catalogs. Other manufacturers have used the expression $L_c t_r$, which is lower than that of Eq. (11.11). When using mixer noise figures quoted in advertising brochures or in the literature, one should be aware of how it was determined.

Types of Mixers[8,9] An ideal mixer is one whose output is proportional to the product of the RF echo signal and the local oscillator (LO) signal. The mixer provides two output frequencies that are the sum and difference of the two input frequencies, or $f_{\text{RF}} \pm f_{\text{LO}}$, assuming $f_{\text{RF}} > f_{\text{LO}}$. The difference frequency $f_{\text{RF}} - f_{\text{LO}}$ is the desired IF frequency. The sum frequency $f_{\text{RF}} + f_{\text{LO}}$ is rejected by filtering. There are however, two possible difference frequency signals at the IF when a signal appears at the RF. One is $f_{\text{IF}} = f_{\text{RF}} - f_{\text{LO}}$, assuming the input RF signal is of greater frequency than the LO frequency. The other possible difference frequency occurs when the RF signal is at a lower frequency than the LO frequency such that $f_{\text{IF}} = f_{\text{LO}} - f_{\text{RF}}$. If one of these is at the desired signal frequency, the other is the *image frequency*. Signals and receiver noise that appear at the image frequency need to be rejected using either an RF filter or an image-reject mixer described later in this subsection.

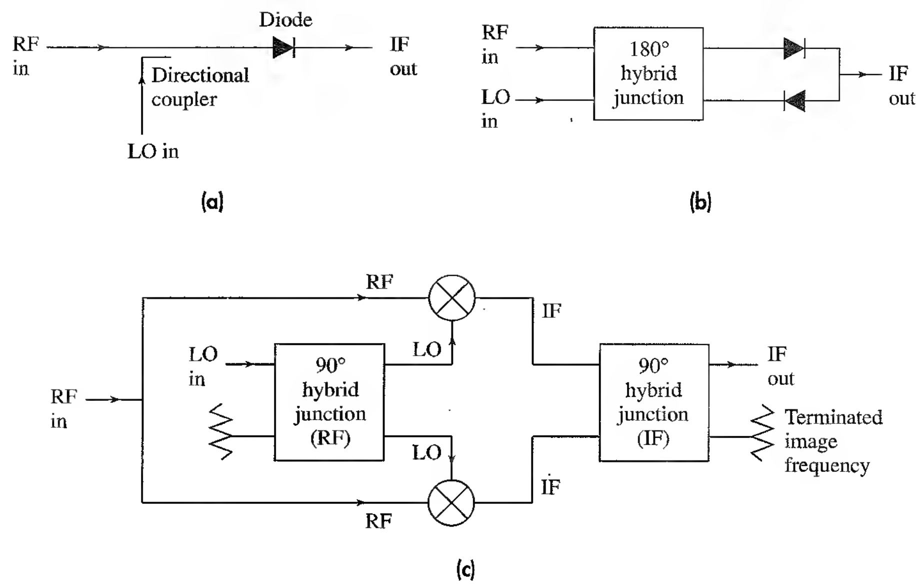
A relatively simple mixer is the *single-ended mixer*, which uses a single diode, as in Fig. 11.2a. The diode terminates a transmission line and the LO is inserted via a directional coupler. A low-pass filter, not shown, following the diode allows the IF to pass while rejecting the RF and LO signals. In a single-ended mixer the image frequency is short-circuited or open-circuited so as to avoid having the noise from the image frequency affect the mixer output.

The diode of a mixer is a nonlinear device and, in theory, can produce intermodulation products at other frequencies, called *spurious responses*. These occur for any RF signal that satisfies the relation[10]

$$mf_{\text{RF}} + nf_{\text{LO}} = f_{\text{IF}} \qquad [11.12]$$

where m and n are integers such that $m,n = \ldots, -2, -1, 0, 1, 2, \ldots$. These are unwanted since they appear within the radar receiver bandwidth. Spurious responses that are

Figure 11.2 Types of mixers: (a) single-ended mixer, (b) balanced mixer, (c) image-rejection mixer.



IF outputs due to the action of the mixer should not be confused with spurious signals, or spurs, that are due to the LO or the receiver power supply and can occur even in the absence of an RF signal. Taylor[7] describes the so-called *mixer chart*, which allows one to determine the combinations of the RF and LO frequencies that are free of strong spurious components. Such a chart indicates the bandwidth available for the mixer as a function of the ratio of the RF and LO frequencies. Taylor points out that the nature of the spurious responses are such that single-conversion receivers generally provide better suppression of spurious responses than double-conversion receivers. The third-order intermodulation product generally affects the dynamic range of the receiver, and is mentioned later under the discussion of dynamic range. There also can be other spurious, or intermodulation, responses from a mixer when two or more RF signals are present at the mixer's input and produce responses within the IF bandwidth.

Noise that accompanies the local oscillator (LO) signal in a single-ended mixer can appear at the IF frequency because of the nonlinear action of the mixer. This noise can be eliminated by inserting a narrowband RF filter between the LO and the mixer. It also has to be a tunable filter if the LO frequency is also tunable. A method to eliminate LO noise that doesn't have these disadvantages is a *balanced mixer*. The balanced mixer also can remove much of the mixer intermodulation products.

A diagram of a balanced mixer is shown in Fig. 11.2b. It can be thought of as two single-ended mixers in parallel and 180° out of phase. At the left of the figure is a four-port junction such as a magic-T, hybrid junction, 3-dB coupler, or equivalent. (Either a 90° or a 180° hybrid can be employed; here it is 180°.) In Fig. 11.2b the LO is applied to one port and the RF is applied to a second port. The signals inserted at these two ports appear in the third port as their sum and in the four port as their difference. A diode mixer is at the output of each of the other two ports. The hybrid junction has the property that

the sum of the RF and LO signals appears at a port containing one of the diode mixers, and at the other port the difference of the RF and LO appears at the diode. The two diode mixers should have identical characteristics and be well matched. The IF signal is obtained by subtracting the outputs of the two diode mixers. In Fig. 11.2b, the balanced diodes are shown reversed so that the IF outputs can be added to obtain the required difference between the two channels. Local-oscillator AM noise at the two diode mixers will be in phase and will be canceled at the output. This mixer configuration also suppresses the even harmonics of both the RF and LO signals.

A *double-balanced mixer* (not shown) utilizes four diodes in a ring, or bridge, network to reduce the LO reflections and noise at the RF and IF ports, achieve better isolation between the RF and LO ports, reject spurious responses and certain intermodulation products, provide good suppression of the even harmonics of both the RF and LO signals, and permit wide bandwidth.[11]

In an *image-rejection mixer*, Fig. 11.2c, the RF signal is split and fed to the two mixers. The LO is fed into one port of a 90° hybrid junction that produces a 90° phase difference between the LO inputs to the two mixers. On the right is an IF hybrid junction that imparts another 90° phase difference in such a manner that the signal frequency and the image frequency are separated. The port with the image signal can be terminated in a matched load. According to Maas,[12] to reduce the image frequency by 20 dB requires that the phase error of the image-rejection mixer be less than 10° and the gain imbalance to be less than 1 dB. Dixon states[13] that the image-rejection mixer provides only about 30 dB of image rejection, which might not be sufficient for some applications. The image-rejection mixer is capable of wide bandwidth, and is restricted only by the frequency sensitivity of the structure of the microwave circuit. It is attractive because of its high dynamic range, good VSWR, low intermodulation products, and less susceptibility to burnout. The noise figure of the image-rejection mixer as well as the balanced mixer will be higher than that of a single-ended mixer because of the loss associated with the hybrid junctions.

The *image-recovery mixer* is an image-rejection mixer designed to reduce the mixer conversion loss by properly terminating the diode in a reactance at the image frequency. Sometimes the lower conversion loss is offset by an increase in noise temperature, a mismatch at the IF, and higher intermodulation products. The improvement using image enhancement is about 1 or 2 dB; hence, the mixer needs to be of low loss so as not to negate the benefit.[14]

Dynamic Range There seems to be no unique definition for the *dynamic range* of a radar receiver. It can generally be described as the ratio of the maximum input signal power to the minimum input signal power the receiver can handle without degradation in performance. "Degradation in performance," however, is not easy to define since it depends on the application. The minimum signal is sometimes taken to be the receiver rms noise level, which depends on the receiver bandwidth. The *minimum detectable signal* S_{min} might be selected as the minimum in the definition of dynamic range; but it also depends on bandwidth and apparently it is seldom used for this purpose.

The maximum signal might be the signal that causes the receiver to saturate (the output no longer increases with an increase in input). Saturation, however, is gradual as the

signal increases in power, so the signal level that results in saturation is not precise. The maximum signal is more usually defined by the acceptable amount of gain compression, which is the deviation of the gain curve (output vs. input) from a straight line. The signal that causes a gain compression of one dB is commonly used for defining the maximum signal. Another criterion for the maximum signal power is based on the onset of *intermodulation distortion.* Intermodulation distortion generally occurs with large signals in the later stages of the receiver. The mixer generates a unique form of intermodulation product called *spurious responses.* It occurs when a harmonic of the local oscillator frequency mixes with a harmonic of the RF signal frequency to create difference frequencies that appear within the IF bandwidth. *Third-order intermodulation* occurs when two equal-amplitude signals within the receiver pass-band at two different frequencies f_1 and f_2 are input to the receiver and produce at the output the frequencies $2f_1 - f_2$ and $2f_2 - f_1$. The maximum signal power is then specified by the third-order intermodulation that can be tolerated when two signals are present within the pass band. The *third-order intermodulation* product is difficult to eliminate by filtering when the two frequencies are close to one another. Another possible indication of the maximum signal is when the echo signal at the mixer approaches the power level of the local oscillator. (The local oscillator power should be at least 7 dB greater than the largest received signal.[15]) No matter what definition is used, the dynamic range is almost always expressed in dB.

A large dynamic range is important if receiver saturation is to be avoided. Once the receiver saturates it can take a finite time to recover before targets again can be detected. Furthermore, when clutter is large enough to saturate the receiver, the MTI improvement factor will be reduced (Sec. 3.7). Saturation of the receiver by clutter echoes causes weak target echoes to be suppressed and not detected even though there might have been adequate improvement factor otherwise. In high-performance radars that must detect small moving targets in the presence of large clutter echoes by doppler processing, the receiver dynamic range must be at least equal to the required improvement factor.

As an example of the variation of the target echo signal power that might be experienced by a radar receiver, assume that an air-surveillance radar has to detect aircraft at ranges from 4 to 200 nmi. This corresponds to a variation in signal power of $(200/4)^4$, which is 68 dB. The average cross section of aircraft might vary from 2 to 100 m^2 (a variation of 17 dB), and the fluctuations in cross section might range over 30 dB. Adding all three factors, the variation of the total target-echo signal might be 115 dB, more or less. This might be an extreme value, but radars that have to detect low cross-section targets could require even greater dynamic range.

Clutter echoes might vary over a range from 60 to 70 dB, or more. The use of STC (sensitivity time control), where the receiver gain is made to vary with time (Sec. 7.8), can reduce the variation in the target echo signal as well as the clutter echo signal. Not all radars, however, can employ STC. Pulse doppler radars, for example, cannot.

The mixer stage is often the limiting factor in dynamic range. A radar receiver that uses a doppler filter bank will have a higher dynamic range because of the narrower bandwidth of each filter. Pulse compression can also increase the dynamic range in proportion to the pulse compression ratio, if the clutter seen by the time sidelobes is not too large. The wider the bandwidth of the receiver (the IF stage) the less will be the dynamic range because of the greater likelihood that mixer intermodulation products (spurious responses)

will be within the frequency band to limit the maximum signal that can be received. A wide bandwidth, as mentioned, also increases the noise level, which reduces dynamic range.

Large dynamic range may be obtained in some radar applications by inserting variable attenuation into the receiver as needed to keep the receiver from being overloaded, but this solution is limited to situations where rapid changes in the input signal are not expected.

Flicker Noise, or 1/f Noise There exists in semiconductors a noise mechanism whose spectral density is inversely proportional to the frequency. It is called *flicker noise* or *1/f noise*,[16] and can be of importance at the lower frequencies. It is quite different from thermal or shot noise, which are independent of frequency. Flicker noise occurs in semiconductor devices such as diodes or transistors, and also in vacuum tubes with oxide-coated cathodes. The frequency relationship of flicker noise is more like $1/f^{\alpha}$, where α varies between 0.8 and 1.3, but it is more common to characterize it as $\alpha \approx 1$.[17] This relationship holds for very low frequencies, lower than might be of practical interest for radar.

The $1/f$ noise is not important for radar receivers whose IF frequencies are greater than a few hundred kilohertz. This is the case for most radar IF frequencies. It can be a factor limiting sensitivity, however, in radars that employ a *homodyne receiver*, also known as a superheterodyne receiver with zero IF. Homodyne receivers are sometimes used in CW radars because of their simplicity. The decrease in sensitivity due to $1/f$ noise at low frequencies might be tolerated for very short-range systems; but when maximum performance is necessary, the effect of the $1/f$ noise can be avoided by use of a superheterodyne receiver with an IF frequency where $1/f$ noise is low.

Oscillator Stability In conventional pulse radars that do not perform doppler processing, stability of the local oscillator, or LO, cannot be ignored but it is usually not a major concern. However, when doppler processing is used to detect moving targets in clutter, as in the MTI radar, the LO has to be quite stable in order to reliably detect the doppler shift. This is why the LO in an MTI radar is called a *stalo*, or stable local oscillator. The MTI improvement factor that can be achieved with a magnetron oscillator transmitter is limited to modest values, so that the demands on oscillator stability can readily be met when a magnetron is used. Power amplifiers such as the klystron, TWT, and the transistor, however, allow much larger improvement factors than a magnetron. Thus greater demands are placed on the stability of the stalos used in such radars. Some high-prf pulse doppler radars that have to detect small targets in the midst of large clutter might encounter clutter echoes that could be 100 dB, or greater, than the target echoes, and thus require highly stable RF sources.

MTI and pulse doppler radars that employ a power amplifier use the sum of the receiver stalo and the coho as the input signal to the power amplifier. (This was indicated in Fig. 3.7 for the MTI radar.) Since the stalo is at a much higher frequency than the coho, it is the stalo that usually sets the limits on what can be achieved. The stalo can have a greater effect on performance than the power amplifier of the transmitter.[18] Thus we only consider the stalo here.

Phase Noise Instability or phase noise in a stalo can be caused by power supply ripple; mechanical and acoustic vibrations from fans, motors, and cooling systems; or by vibrations of the platform (such as an aircraft or ship); as well as spurious responses and

noise from the stalo itself. Phase noise is usually considered in the frequency domain, but in the time domain it can be thought of as being due to the deviation of the oscillator signal from a perfect sinewave. There is also amplitude-modulation noise associated with oscillators, but AM noise is usually small compared to phase noise. If not, it can be reduced by balanced mixers or other means.

In Sec. 3.7 the effect of equipment instabilities on MTI performance was mentioned. There it was shown that a pulse-to-pulse change in phase, $\Delta\phi$, limits the improvement factor of a two-pulse MTI to $I_f = (\Delta\phi)^{-2}$. When an MTI or pulse doppler radar uses many pulses to perform doppler filtering, this simple expression for improvement factor no longer applies. A different model has to be considered.

The reader is referred to Fig. 3.36 for an example of the spectrum of an oscillator as might be used in a mixer. (There is also further discussion of oscillator stability for doppler radars where this figure appears in Sec. 3.7.) In addition to the narrow spike at d-c due to the carrier (not shown in the figure), there is a noise spectrum that decreases monotonically with increasing frequency. There are also spikes, or *spurs*, that are usually caused by the power supply or vibrations. At the higher frequencies the phase noise levels off and is characterized by a uniform noise floor. The ordinate in Fig. 3.36 is the noise power within a one-hertz bandwidth relative to that of the carrier. It should be multiplied by the receiver bandwidth to obtain the actual power at the receiver.

Although Fig. 3.36 might be the noise from a stalo, it also represents the noise radiated by the transmitter since the stalo is a major part of the signal that excites the power amplifier transmitter. This noise may seem far down from the peak of the carrier, but the spectrum of the echo from stationary clutter is the same as the spectrum of the transmitter. (Internal motion of the clutter can further increase the spectrum of the received echo signal.) As mentioned, clutter can be quite large compared to the weak moving target echo. The MTI or pulse doppler radar may be able to attenuate the main clutter line at d-c, but the clutter spectrum often has components at frequencies where doppler-shifted echoes from moving targets are expected. These components can mask the desired target echoes. Good performance of an MTI or a pulse doppler radar requires that the transmitter spectrum, and the clutter-echo spectrum it produces, be low enough to detect the slowly moving weak targets that are of interest. It would not be unusual to find that oscillator noise can be the limiting factor in some high-performance radars that must detect low-speed, low cross-section moving targets in heavy clutter. Good oscillator design is therefore important for achieving good radar performance.

The effect of phase noise can be determined by measuring the phase-modulation spectrum of the stalo and using it to obtain the MTI improvement factor. The procedure will not be given here. It is outlined by Taylor[18] and given in more detail with examples by Goldman.[19] Since the stalo is part of the transmitter as well as part of the receiver, the effect of phase noise on MTI performance will be range dependent. At the shorter ranges, or time delays, greater stalo noise can be tolerated at frequencies closer to the carrier (lower target doppler frequency shifts) than at longer ranges. For this reason, the effect of stalo stability needs to be computed for several ranges.

Oscillator phase noise can be a serious limitation to the performance of a modern high-performance MTI or a pulse doppler radar. Its effect has to be found with measurements and analysis more complicated than was indicated in Sec. 3.7.

Types of Stable Oscillators[20,21] Almost all of the oscillators used for stable sources can be thought of as consisting of an amplifier, a resonant circuit that determines the frequency and the phase noise, and feedback to generate oscillation. The amplifier is often a transistor. The following is a brief listing of the various oscillators that have been considered for use as stable sources.

Crystal Oscillator The mechanically vibrating piezoelectric quartz crystal has been an important device for producing stable oscillators ever since the early days of commercial radio.[22] A piezoelectric material is one which mechanically deforms along one crystal axis when an electric potential is applied along another axis. Conversely an electrical potential is obtained when a mechanical deformation occurs. The piezoelectric crystal is used as the resonator in the feedback circuit of a transistor oscillator. It is often mounted in a small-size temperature-controlled oven and isolated from vibrations. It is a very stable source at low frequencies (10 to 180 MHz), but its output can be multiplied in frequency to provide stable signals in the microwave region.

Frequency Multiplier A low-frequency stable oscillator can be multiplied to a higher frequency by applying its signal to a nonlinear device such as a diode or varactor to generate harmonics of the fundamental frequency. A filter is used to select the desired harmonic. The phase noise power, however, increases as the square of the frequency-multiplication ratio. For example, when a 10-MHz stable source is multiplied to 10 GHz, its noise is increased by 60 dB. In addition, there can also be additive phase noise produced in frequency multipliers. In spite of the increase in noise with multiplication in frequency, multiplication is a good method for taking advantage of the excellent stability of a low-frequency sources to obtain a stable oscillator at radar frequencies.

Dielectric Resonator Oscillator (DRO) The resonant circuit in this type of oscillator is a dielectric material such as a sapphire crystal, ceramic,[23] or titanate in a regular geometric form that acts as a microwave resonant cavity. The high dielectric constant of the resonator allows it to be much smaller in size compared to a metallic cavity resonator. It is among the most stable of room-temperature oscillators. Because of its small size it has a relatively high Q and may be quite rigid so as to reduce its sensitivity to shock and vibration. When the dielectric resonator is made larger to obtain even higher values of Q and enhanced frequency stability, it might be more sensitive to temperature changes and vibration. The DRO has been a popular device for application as a low-noise, stable oscillator at microwave frequencies.

SAW Oscillator The surface acoustic wave (SAW) device can also be used as the resonator in a feedback oscillator. SAW oscillators can be quite small and can be obtained from about 100 MHz to 3 GHz. Ewell[20] states that the phase noise of a SAW oscillator can be worse than that of a frequency-multiplied crystal oscillator at low offset frequencies (1 kHz for example) from the carrier, but it can be better at high frequency offsets (greater than 10 kHz).

YIG Oscillator A small sphere of yttrium iron garnet (YIG) suspended within a resonant cavity with an applied magnetic field can act as the resonant device of an oscillator. The resonant frequency of a spherical YIG crystal depends only on the applied magnetic field and not on its dimensions. It has a relatively high level of phase noise, but it has the advantage of being tunable by changing the applied magnetic field.

Klystron Oscillator and Gunn Oscillator The reflex klystron oscillator (originally used as the local oscillator of many a World War II radar receiver) and the Gunn diode oscillator are two very different type of devices. Both have relatively high phase noise, but when coupled to a high-Q external resonant cavity they can be of high stability. The use of superconducting high-Q cavities can produce "extremely low phase noise levels."[20]

High-Temperature Superconducting Oscillators As was mentioned in Sec. 3.7, the phase noise of oscillators can be improved by employing very low loss superconductive resonators, especially those that are superconductive at the temperature of liquid nitrogen, 77 degrees.[24]

Direct Digital Synthesis[25] A *frequency synthesizer* produces one or more frequencies over a wide spectrum by translating the stable frequency of a precision frequency source, such as a crystal-controlled oscillator. In *direct synthesis* a single precision oscillator is multiplied and/or divided to obtain a desired frequency. When this process is performed digitally, it is called *direct digital synthesis* (DDS).[26] A DDS can generate the multiple frequencies needed for the stalo, coho, a second LO if dual conversion is used, and timing frequencies. It can also provide linear or nonlinear FM for pulse compression systems. A DDS generally uses a phase accumulator (to establish the time sequence), sine lookup table (to establish the amplitude of the signal waveform), a digital-to-analog converter, low pass filter, and frequency multiplier or heterodyne to translate to a higher frequency.[27] It has the advantage of extremely fast frequency switching, small size steps in frequency, excellent phase noise, reasonably good spurious performance, transient-free (phase continuous) changes in frequency, flexibility in applying modulation, and it achieves its good performance in a small volume.

A/D Converters The A/D converter, which changes analog signals to digital signals, is an important component of digital processing. There are many different ways it has been implemented.[28] Its performance for radar is judged by the number of bits into which it can quantize a signal and the sampling rate at which it can operate. As was mentioned in Sec. 3.5 (where the effect of the A/D converter on MTI performance was discussed and some examples of performance were given), the number of bits into which the A/D converter can quantize a signal decreases as the sampling rate, or bandwidth, increases. Thus the larger the bandwidth of the signal the more difficult it is to maintain good performance. The A/D converter sometimes can be a limitation in wideband radar or when large clutter attenuation is required.

Bandpass Sampling at IF Digital signal processing that is conducted at baseband (video) requires two baseband A/D converters and an in-phase and a quadrature channel. Although

baseband digital processing has been widely used, there are limitations. The two baseband converters have to be well balanced over a wide dynamic range and there cannot be significant phase errors between the two channels (the phase difference between the two channels cannot differ significantly from 90°). Waters and Jarrett[29] indicate that these problems do not appear if the A/D conversion is performed in the bandpass portion of the receiver at IF. The in-phase and quadrature components are obtained by a single A/D converter from the samples taken directly from the original IF signal. The phase errors between the two channels are considerably smaller with IF sampling than with baseband sampling. Although only one channel is needed in bandpass sampling, its sampling rate has to be greater than that of the A/D converters used in baseband sampling. Further discussion can be found in Sec. 3.5.

Digital Radar Receiver There does not seem to be a unique, well-accepted definition of a digital receiver. A digital radar receiver, ideally, could be thought of as one that is completely digital with a wide dynamic range A/D converter that operates directly on the signal received at the antenna terminals. This would be followed by a highly capable computer to perform the functions found in a radar receiver. It is difficult, however, to achieve such a receiver with the bandwidth and large dynamic range required for high-performance microwave MTI and pulse doppler radars. More realistically, a digital radar receiver might be one that uses an analog RF amplifier and mixer, and even analog IF circuitry, followed by an IF A/D converter and digital video processing.

A different and more practical definition of a digital radar receiver was proposed by Wu and Li,[30] who stipulate that such a receiver have two significant differences compared to analog radar receivers. It should utilize (1) a direct digital synthesizer (DDS) as the local oscillator and (2) direct bandpass sampling at IF before detection, with all subsequent processing being done digitally.

In addition to a high-speed A/D converter with many bits of quantization, a digital receiver requires the digital processing to have sufficient speed to operate in real time and to have a large enough information storage memory. There is little doubt that the major advances in radar and its increased applicability since the 1970s have been due to the phenomenal advances in digital processing technology. It is likely that "digits" will continue to be a major driver of future advances in radar performance.

Russian Cyclotron Wave Electrostatic Amplifier[31–33] Solid-state amplifiers have been a popular choice for the front-end of a radar receiver, but they are not the only choice. A Russian receiver development, called the cyclotron wave electrostatic amplifier (CWESA), has been a popular receiver for certain types of radars because of characteristics not available with other devices. It is also more usually known as an *electrostatic amplifier* (ESA). The ESA is said to have a low noise figure, bandwidths of 5 to 10 percent, linear phase variation with frequency, and other attributes suitable for a receiver front end; but its uniqueness is that it can sustain a high level of input power without additional protection and it can recover quickly from overload. Duplexers or receiver protectors are not needed.

In the ESA an electrostatic cyclotron wave is launched on a thin electron beam in an input structure; it is amplified in an intermediate structure, and then coupled to an output

structure. The thin electron beam at the cathode might have dimensions of 0.03 by 0.7 mm and a current of 250 to 280 μA. The theory[31] of this device will not be summarized here except to say that a longitudinal magnetic field is required so that the input signal, when coupled to the electron beam, results in cyclotron motion of the electrons. Permanent magnets are used to reduce weight. At *S* or *C* bands, these units are said to weigh approximately 2 kg, have an approximate volume of one liter, and a power consumption of 1 to 1.5 W. A 1.0 dB noise figure was achieved at frequencies up to 3 GHz and a 2.4 dB noise figure at 10 GHz.

When a large signal appears at the input to the ESA, it causes a large reflection coefficient (a large VSWR) so that the signal is entirely reflected and is not absorbed, which is unlike diode receiver protectors that absorb the input energy. Thus these receivers have been used in radars without additional duplexers or diode receiver protectors. When the overload is removed the device quickly returns to service, typically in about 20 ns at frequencies above *S* band.[31,32] Longer values of recovery time are experienced at lower frequencies. It has been claimed that in radar applications the ESA can withstand peak powers of 10 kW and average powers of 300 W at frequencies above *S* band, and higher powers at lower frequencies.

Sometimes a transistor amplifier is added as a second stage to obtain higher gain. Such a combined ESA and transistor, operating from 7 to 7.4 GHz with a single high-voltage supply of 400 V in addition to small filament and transistor amplifier supply voltages, produced a noise figure of 3.4 dB and a gain of approximately 23 dB. It could withstand in excess of 5 kW of peak and 150 W of average power at the input and recover in less than 50 ns. When a transistor second stage is used, such a device sometimes is called an *electrostatic combined amplifier* (ESCA).

A tunable version of the ESA was said to demonstrate very rapid tuning over a 50 percent bandwidth, with an instantaneous bandwidth of 1 percent.

The rapid recovery time of this amplifier makes it attractive for use with pulse doppler radars which require a high prf. Pulse doppler radars operate with high duty cycles so there is little range-space available. Long recovery times reduce the available range-space. If the duty cycle were 10 percent and the pulse width were 1 μs, a diode protector recovery time of 1 μs would significantly increase the receiver dead time and increase the minimum range. The 20 ns recovery time of the ESA would hardly be noticed. The duty cycles of high-prf pulse doppler radars can be as high as 0.3 to 0.5, which makes receiver recovery time even more important.

The testing of a 200-MHz bandwidth *X*-band ESA combined with a transistor second stage as an ESCA receiver for a high-prf pulse doppler radar was described by Ewell.[31] Pulse repetition rates were from 1 kHz to several hundred kHz and pulse durations less than 1 μs. For this application, the ESCA was considered superior to conventional gas TR tubes whose recovery times were too long and too unpredictable. They were also superior to multipactor discharges which had good recovery time but had high spike leakage that required varactor diode receiver protectors. They also were costly and required additional components such as an oxygen generator, ion pump, and cooling system. Ewell's measurements appear to confirm the consistency of this device to meet pulse doppler radar system requirements. It provided protection from overload, fast recovery time, linearity, and electronic control of dynamic range.

Because of its size, the ESA is not suitable for most applications of active-aperture phased array radar; but there are many important radar applications where an active aperture is not necessary. The ESA is attractive in those radar systems where a single or only a few receiver channels are used. At the time of publication of the cited references for this subsection, there were roughly ten thousand of these devices manufactured and in use in a number of systems around the world, mainly in Russia and China. An example is its use in the Russian S300 PMU air defense and anti-tactical ballistic missile (ATBM) system (NATO designation SA-10). This employs an *X*-band space-fed phased array radar with pulse doppler waveform designed to operate in high clutter and electromagnetic countermeasure environments.[31] The ESA is also found in the Russian S300V (NATO designation SA-12) air defense system. Barton points out that the electrostatic amplifier tube helps make the total RF loss in these Russian receiving systems significantly lower than the loss found in comparable Western systems.[34]

Phase Detector, Phase-Sensitive Detector In Fig. 3.7 of Sec. 3.1, the *phase detector* was introduced as the device in an MTI receiver that extracted the doppler frequency shift of an echo signal. It compared the echo signal to a reference signal (the coho) which was coherent with the MTI transmitter signal. In the MTI phase detector, it is the rate of change of phase of the echo signal with time that is of interest since it determines the doppler frequency shift of the echo from a moving target. In Fig. 4.4 of Sec. 4.2, the *phase-sensitive detector* was shown in the amplitude-comparison monopulse tracking radar to allow the extraction of the sign of the angle error along with its magnitude. The input to this detector was the angle-error signal and the signal from the sum channel which acted as the reference. In both the MTI radar and the monopulse tracker two sinusoidal voltage inputs were available to a nonlinear device. The two were coherent with respect to one another in that they could be thought of as being from the same source. In both these detectors, one of the two voltages is the reference and the other is the received echo signal.

Taylor[35] points out that the distinction between a phase detector and a phase-sensitive detector is not always clear because of the similarity of the analog circuits that perform these two functions. He states that it is generally agreed that a phase detector is one in which only phase information is present in the output; a phase-sensitive detector is one in which both phase and amplitude information are in the output; and a mixer when phase, amplitude, and frequency information are present in the output. He also points out that "doppler frequency shifts are excepted in this convention."

Krishnam[36] indicates that the difference between these two detectors is in the actual operating conditions and not the hardware. He states that it had been usual to assume that the reference and the signal are of the same amplitude for the phase detector. For the phase-sensitive detector it was common to assume that the reference is much larger than the signal. He then shows that other assumptions can be made. He denotes $V_1 = E_1 \sin \omega t$ as the reference and $V_2 = E_2 \sin(\omega t + \phi)$ as the signal. For his particular detector model, he then shows that when E_1 is exactly equal to E_2, the output is $E_0/E_1 = 2(|\cos \phi/2| - |\sin \phi/2|)$, which is approximately linear with respect to ϕ over the range $0 < \phi < \pi/2$. Under these conditions, the device can be considered as a phase detector. When $E_2 >> E_1$ (signal is large compared to the reference), the output is $E_0 = 2E_1 \cos \phi$, which also is a phase detector. When $E_2 < E_1$ so that the reference is larger than the largest

signal E_2 , the device is shown to be a "perfectly linear" phase-sensitive detector with an output $E_0 = +2E_2$ when $\phi = 0$, and $E_0 = -2E_2$ when $\phi = \pi$.

Example of a Receiver One seldom finds in the radar literature a paper on the design of a radar receiver. For some reason, receiver designers do not prepare such papers, or the journal editors and referees do not accept them. There is, however, at least one paper of which I am aware that describes the receiver for the original Aegis AN/SPY-1A shipboard air-defense system.[37] The receiver is in two parts. One part is the on-array portion which contains the low-noise amplifiers and related components. It is mounted to the rear of each of the four antenna faces of Aegis to minimize pre-RF amplifier losses. The other part is located both in the fore and aft deckhouses and contains components with minimal impact on the noise figure. There are eleven receiver channels: three monopulse tracking channels, one sidelobe interference blanker channel, six auxiliary ECCM sidelobe canceler channels, and one auxiliary channel that acts as a spare. Each channel has two inputs so they can be time-shared between two antenna arrays to reduce cost.

There is too much in the paper to adequately summarize here, but it is recommended as being one of the few examples available that provides an overview of radar receiver engineering not usually found in radar texts.

11.4 DUPLEXERS AND RECEIVER PROTECTORS

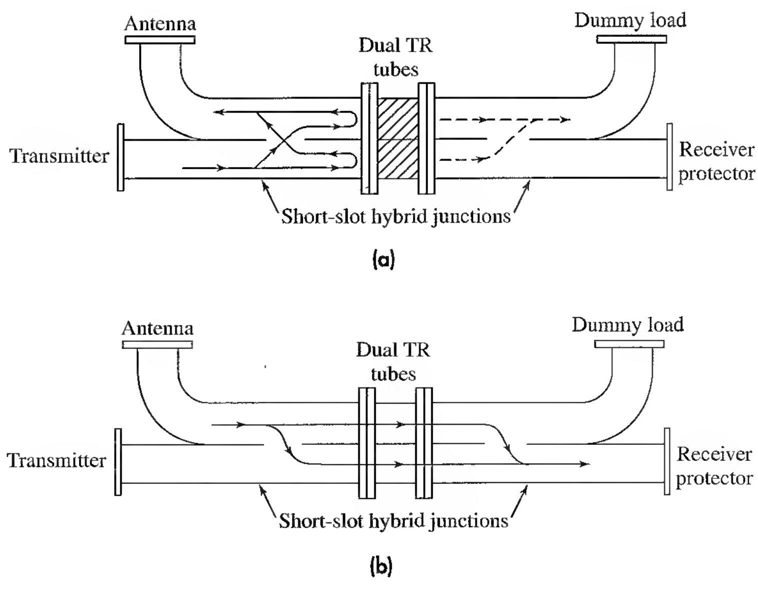
A pulse radar can time share a single antenna between the transmitter and receiver by employing a fast-acting switching device called a *duplexer.* On transmission the duplexer must protect the receiver from damage or burnout, and on reception it must channel the echo signal to the receiver and not to the transmitter. Furthermore it must accomplish the switching rapidly, in microseconds or nanoseconds, and it should be of low loss. For high-power applications, the duplexer is a gas-discharge device called a TR (transmit-receive) switch. The high-power pulse from the transmitter causes the gas-discharge device to break down and short circuit the receiver to protect it from damage. On receive, the RF circuitry of the "cold" duplexer directs the echo signal to the receiver rather than the transmitter. Solid-state devices have also been used in duplexers. In a typical duplexer application, the transmitter peak power might be a megawatt or more, and the maximum safe power that can be tolerated by the receiver might be less than a watt. The duplexer, therefore, must provide more than 60 to 70 dB of isolation between the transmitter and recovery with negligible loss on transmit and receive.

The duplexer cannot always do the entire job of protecting the receiver. In addition to the gaseous TR switch, a receiver might require diode or ferrite limiters to limit the amount of leakage that gets by the TR switch. These limiters, which have been called *receiver protectors*, also provide protection from the high-power radiation of other radars that might enter the radar antenna with less power than necessary to activate the duplexer, but with greater power than can be safely handled by the receiver. There might also be a mechanically actuated shutter to short-circuit and protect the receiver whenever the radar is not operating. Sometime the entire package of devices has been known as a *receiver*

protector.[38] The term is ambiguous, since receiver protector is also the name for the diode limiter or similar device that follows the duplexer for the purpose of reducing the leakage power passed by the duplexer. In this text the term receiver protector is used to denote a limiter that follows the duplexer. The duplexer, receiver protector, and other devices for preventing receiver damage are better known as the *duplexer system*, so as to prevent confusion by the same term (receiver protector) being used to describe the entire receiver protection system as well as one part of it.

Balanced Duplexer The balanced duplexer, shown in Fig. 11.3, is based on the short-slot hybrid junction which consists of two sections of waveguides joined along one of their narrow walls with a slot cut in the common wall to provide coupling between the two.[39] (The short-slot hybrid junction may be thought of as a broadband directional coupler with a coupling ratio of 3 dB.) Two TR tubes are used, one in each section of waveguide. In the transmit condition, Fig. 11.3a, power is divided equally into each waveguide by the first hybrid junction (on the left). Both gas-discharge TR tubes break down and reflect the incident power out the antenna arm as shown. The short-slot hybrid junction has the property that each time power passes through the slot in either direction, its phase is advanced by 90°. The power travels as indicated by the solid lines. Any power that leaks through the TR tubes (shown by the dashed lines) is directed to the arm with the matched dummy load and not to the receiver. In addition to the attenuation provided by the TR tubes, the hybrid junctions provide an additional 20 to 30 dB of isolation.

On reception the TR tubes do not fire and the echo signals pass through the duplexer and into the receiver as shown in Fig. 11.3b. The power splits equally at the first junction and because of the 90° phase advance on passing through the slot, the signal recombines in the receiving arm and not in the arm with the dummy load.



igure 11.3 Balanced duplexer sing dual TR tubes and twa short-slat ybrid junctions. (a) Transmit condition nd (b) receive canditian.

The balanced duplexer is a popular form of duplexer with good power handling capability and wide bandwidth.

TR Tube The TR tube is a gas-discharge device designed to break down and ionize quickly at the onset of high RF power, and to deionize quickly once the power is removed. One construction of a TR consists of a section of waveguide containing one or more resonant filters and two glass-to-metal windows to seal in the gas at low pressure. A noble gas like argon in the TR tube has a low breakdown voltage, and offers good receiver protection and relatively long life. TR tubes filled only with pure argon, however, have relatively long deionization times (long recovery times) and are not suitable for short-range applications. Adding water vapor or a halogen gas to the tube speeds up the deionization time, but such tubes have shorter lifetimes than tubes filled only with a noble gas. Thus a compromise must usually be accepted between fast recovery time and long life.

To insure reliable and rapid breakdown of the TR tube on application of high power, an auxiliary source of electrons is supplied to the tube to help initiate the discharge. This may be accomplished with a "keep-alive," which produces a weak d-c discharge that generates electrons that diffuse into the TR where they assist in triggering the breakdown once RF power is applied by the transmitter. An alternative is to include a small source of radioactivity, such as tritium (a radioactive isotope of hydrogen), which produces low-energy-level beta rays to generate a supply of electrons.[40] The tritium is in compounded form as a tritide film. The radioactive source, sometimes called a *tritiated ignitor*, has the advantage of not increasing the wideband noise level as does a keep-alive discharge (by about 50 K) and has longer life (by an order of magnitude), but it passes more leakage energy so that it requires one or more cascaded PIN diode limiter stages to further attenuate the leakage.[41] The tritium ignitor needs no active voltages, so it allows the receiver protector to function with the radar off without the need for a mechanical shutter to protect the radar from nearby transmissions. Being a radioactive device, however, does cause concern about its handling and disposal. The combination of the tritium-activated TR followed by a diode limiter has been called a *passive TR-limiter*.

The TR is not a perfect switch; some transmitter power always leaks through to the receiver. The envelope of the RF leakage might be similar to that shown in Fig. 11.4. The short-duration, large-amplitude *spike* at the leading edge of the leakage pulse is the result of the finite time required for the TR to ionize and break down. Typically, this time is of the order of 10 nanoseconds. After the gas in the TR tube is ionized, the power leaking through the tube is considerably reduced from the peak value of the spike. This portion

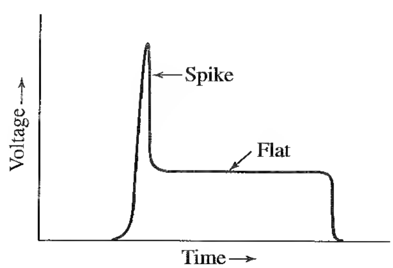


Figure 11.4 Leakage pulse through a TR tube.

of the leakage pulse is called the *flat*. Damage to the receiver front-end may result when either the energy contained within the spike or the power in the flat portion of the pulse is too large. The spike leakage of TR tubes varies with frequency and power and whether or not the tube is primed with electrons, but might be "typically" about one erg. The attenuation of the incident transmitter power might be of the order of 70 to 90 dB.

A fraction of the transmitter power incident on the TR tube is absorbed by the discharge. This is called *arc loss*. It might be 0.5 to 1 dB in tubes with water vapor and 0.1 dB or less with argon filling. On reception, the TR tube introduces an insertion loss of about 0.5 to 1 dB. The life of a TR tube is determined more by the amount of leakage power it allows to pass or when its recovery time becomes excessive rather than by its physical destruction or wear.

Solid-State Receiver Protectors, Diode Limiters Improvements in receiver sensitivity sometimes are obtained with front-ends and mixers that are more sensitive to damage from RF leakage. Such sensitive devices require better protection from the RF leakage of conventional duplexers. A PIN diode limiter placed in front of the receiver helps reduce the leakage and act as a *receiver protector*. A diode limiter passes low power with negligible attenuation, but above some threshold it attenuates the signal so as to maintain the output power constant. This property can be used for the protection of radar receivers in two different implementations depending whether the diodes are operated unbiased (self-actuated) or with a d-c forward-bias current. Unbiased operation without the use of an external current supply is also known as *passive*. It has the advantage of almost unlimited operating life, fast recovery time, no radioactive priming, and versatility to perform multiple roles.[38] Its chief limitation is its low power handling. A passive solid-state limiter for X-band WR90 waveguide with 7 percent bandwidth and a 1-μs pulse width had a peak power capability of 10 kW and a CW power capability of 10 W.[42] Its insertion loss was 0.6 dB, leakage power was 10 mW, and a 1.0-μs recovery time. When used with a 40-μs pulse width, this limiter could withstand 2 kW of peak power and 300 W CW, with a loss of 0.8 dB.

Biasing of the diodes during the high-power pulse, also known as *active*, is capable of handling a great deal more power than when operated passively. The diode is biased into its low impedance mode prior to the onset of the transmitter pulse. Although the active diode-limiter offers many advantages for use with duplexers, it does not protect the receiver when the bias is off. It thus offers poor protection against nearby asynchronous transmissions that arrive at the radar during the interpulse period or when the radar is shut down.

Incorporation of Sensitivity Time Control[43,44] In Sec. 7.8 the use of sensitivity time control (STC) was described as a method to reduce the effects of nearby large clutter echoes without seriously degrading the detection of desired targets at short range. STC is the programmed change of receiver gain with time, or range. At short ranges the receiver gain is lowered to reduce large nearby clutter echoes. As the pulse travels out in range the gain is increased until there are no more clutter echoes.

There are advantages for having the STC in the RF portion of the radar just ahead of the receiver. STC can be applied by biasing the diodes of a receiver protector to provide

a time-varying attenuation without adding to the receiver noise figure. There is no increase in insertion loss to obtain the STC action, above that inherent in the design of the receiver protector. The PIN diode stages provide the self-limiting action during transmit and the STC function during receive. The nonlinear nature of the diode requires a linearizing circuit to achieve the desired variation of attenuation with time. The STC variation depends on the nature of the terrain seen by the radar. A digitally controlled STC drive with random access memory allows the radar designer to employ different STC response profiles according to the various types of terrain that might be seen by the radar.

Varactor Receiver Protectors With fast-rise-time, high-power RF sources, the receiver protector may be required to self-limit in less than one nanosecond. This can be achieved with fast-acting PN (varactor) diodes. A number of diode stages, preceded by plasma limiters, might be employed. In one design, an *X*-band passive receiver protector was capable of limiting 1-ns rise time, multikilowatt RF pulses to 1-W spike levels.[45]

Ferrite Limiters The ferrite limiter has very fast recovery time (can be as low as several tens of nanoseconds), and if the power rating is not exceeded, it should have long life. The spike and flat leakage are low and it has been able to support a peak power of 100 kW;[46] but the insertion loss is usually higher (1.5 dB) and the package is generally longer, heavier, and more expensive than other receiver protectors. Except for the initial spike, the ferrite limiter is an absorptive device rather than a reflective device (as is a gas-tube TR) so that the average power capability of these devices can be a problem. Air or liquid cooling might be required. A diode limiter usually follows the ferrite limiter to reduce the leakage at high peak power.

Pre-TR Limiter[38,47] A pre-TR is a gaseous tube placed in front of a solid-state limiter. The function of the pre-TR is to reduce the power that has to be handled by the diode limiter. (It is similar to what was called a passive TR-limiter earlier in this section.) The pre-TR gas tube has high power handling capability, can operate with long pulses, has very fast recovery time, contains a radioactive priming source, but has limited operating life. Very high average power levels may require liquid cooling of the pre-TR mount. End of life for a pre-TR tube usually is caused by the increased recovery times that result from the cleanup of the gas within the tube.

The pre-TR tube can be a quartz cylinder filled with chlorine or a mixture of chlorine and an inert gas. Chlorine, a halogen gas, has a very rapid recovery time; typically a fraction of a microsecond for pulse widths up to 10 μs. The tube is mounted in a waveguide iris. In some cases, the quartz pre-TR tube can be designed to be field replaceable once it reaches end of life.

Mutipactor[46,48] The recovery times of high-power duplexers discussed thus far are from a fraction of a microsecond to several tens of microseconds. By employing the principle of multipacting, a recovery time as short as 5 or 10 ns is possible. Fast recovery time is important for high prf and high duty cycle radars. The multipactor is a vacuum tube and does not have the long recovery characteristics of a gas-filled tube. It contains surfaces capable of large secondary electron emission upon impact by electrons. The secondary

emission surfaces are biased with a d-c potential. The presence of RF energy causes electrons to make multiple impacts that generates by secondary emission a large electron cloud. The electron cloud moves in phase with the oscillations of the applied RF electric field to absorb energy from the RF field. RF power is dissipated thermally at the secondary emission surfaces, and the device requires liquid cooling to remove the absorbed power. Since it is a vacuum device, the recovery time of the multipactor is extremely fast. The flat-leakage power passed by the multipactor is often high enough to require a passive diode limiter to follow it. The multipactor offers no protection when the power is turned off. It has the disadvantage of being complex in that it requires liquid cooling, an ignitor electrode to ensure that multipacting starts quickly, an oxygen source to maintain the magnesium oxide surface that provides the secondary emission electrons, and a pump to maintain a good vacuum.

Solid-State Duplexers There has always been a desire to replace the gas-discharge duplexer with an all-solid-state duplexer because of the potential for long life, fast recovery time, no radioactive priming, and versatility. Although passive operation is desired, it is limited in power. The lowest loss and highest power handling are obtained with active circuits in which the PIN diodes are switched in synchronism with the transmitter pulses. Generally, diodes that can handle high power will have longer recovery times and tend to have higher leakage power—so that they might require additional stages of lower level limiters with increased loss and increased cost. A failure of the active drive circuit, however, could cause destruction of the diode switches as well as the receiver.

Several examples of all solid-state duplexers have been described in the literature. An L-band self-switching duplexer design used four PIN diodes that were biased by four fast-acting decoupled varactor detector diodes.[49] These detector diodes bias the PIN diodes into conduction in a time considerably shorter than the rise time of the RF power pulse. The device could handle 100-kW peak power with 100-W average power and a 3-μs pulse width. Its insertion loss was 0.5 dB. The duplexer was followed by a low-power multiple stage varactor limiter that reduced the spike and flat leakage of 2.8 kW and 32 W peak respectively to levels low enough that low-noise amplifiers were adequately protected. The recovery time was about 15 μs. A UHF solid-state duplexer also using four diodes was reported to have 300-kW peak power, 5-kW average power, 60-μs pulse width, and an insertion loss of 0.75 dB. A C-band solid-state duplexer with 16 PIN diodes was capable of 1-MW peak power with a 14-μs pulse width, 0.01 duty cycle, and insertion loss of less than 1 dB.[50] This device was followed by an additional low-power diode switch with an insertion loss of about 0.6 dB. It provided an isolation of 60 dB, making the total isolation of the duplexer system over 100 dB.

Circulators as Duplexers The ferrite circulator is a three- or four-port device that can, in principle, offer isolation of the transmitter and receiver. In the three-port circulator, the transmitter may be connected to port 1. It radiates out of the antenna connected to port 2. The received echo signal from the antenna is directed to port 3 which connects to the receiver. The isolation between the various ports might be from 20 to 30 dB, but the limitation in isolation is determined by the reflection (due to impedance mismatch) of the transmitter signal from the antenna that is then returned directly to the receiver. For

Table 11.1 Comparison of various types of duplexing devices

Device	Recovery Time	Average Power	Peak Power
TR tube	<1 μs to 100 μs		1 MW
Pre-TR	50 ns to 1 μs	50 kW	5 MW
Diode limiter	50 ns to 10 μs	1 kW	100 kW
Ferrite limiter	20 ns to 120 ns	10 W	100 kW
Multipactor	1 ns to 20 ns	500 W	80 kW
Electrostatic amplifier	20 ns	300 W, or higher	10 kW, perhaps as high as 500 kW

example, if the VSWR (voltage standing wave ratio) of the antenna were 1.5 (a pretty good value), about 4 percent of the transmitter power will be reflected by the antenna and return to the receiver. This corresponds to an isolation of 14 dB. If the VSWR were 2.0, the effective isolation is only 10 dB. To limit damage, a good receiver protector needs to be included. Circulators can be made to withstand high peak and average power; but large power capability generally comes with large size and weight. For example, an *S*-band differential phase-shift waveguide circulator that weighs 80 pounds has essentially the same insertion loss, isolation, and bandwidth of an *S*-band miniature coaxial Y-junction circulator that weighs 1.5 oz.[51] The larger circulator, however, can handle 50 kW of average power while the smaller circulator is rated at 50 W. (The ratio of powers exceeds the ratio of weights.)

Small-size circulators, usually in conjunction with a receiver protector, often are used as the duplexer in solid-state TR modules for active aperture phased arrays. (Note that, unfortunately, the term "TR" has been used for both T/R modules and TR duplexer gas-discharge tubes.)

Summary of Performance Table 11.1 summarizes the performance of various duplexer devices in term of recovery time and power handling. This table is adapted from the paper by Bilotta,[38] but modified from information in other references cited previously. The values in this table depend on frequency and other factors so they should not be considered as absolute limits, but only as an approximate guide.

11.5 RADAR DISPLAYS

Originally the radar display had the important purpose of visually presenting the output of the radar receiver in a form such that an operator could readily and accurately detcct the presence of a target and extract information about its location. The display had to be designed so as not to degrade the radar information and to make it easy for the operator to perform with effectiveness the detection and information extraction function. It was not uncommon for an operator to employ a grease pencil to mark on the face of a cathode-ray-tube display the location of a target from scan to scan and manually extract the

target speed and direction. As digital signal processing and digital data processing improved, more and more of the detection and information extraction process was performed automatically by electronic means so that the role of the operator was less. Processed detection and target information now are displayed to the operator who has little responsibility for making the actual detection decision. Instead of displaying only detections, many surveillance radars display target track vectors along with auxiliary alphanumeric information to an operator.

When the display is connected directly to the output of the radar receiver without further processing, the output is called *raw video.* When the receiver output is first processed by an automatic detector or an automatic detector and tracker before display, it is called *synthetic video* or *processed video.* The requirements for the display differ somewhat depending whether raw or processed video is displayed. Some radar operators prefer to see on a display the raw video lightly superimposed on the processed video.

In many cases the operator does not see the unprocessed output of the radar. An example is the Nexrad doppler weather radar in which the radar measures three parameters in each resolution cell: the amplitude of the echo signal (proportional to its radar cross section), the mean radial velocity (from the doppler frequency shift) of the meteorological scatterers, and the variance of the radial velocity (a measure of the motion of the individual scatterers within the resolution cell). These three meteorological echo parameters in each resolution cell are passed to a computer that generates the many different types of weather products such as maps of precipitation, wind shear at various horizontal and vertical planes, mesocyclones, tornadoes, prediction of flooding, and many others.

The radar display is now more like the familiar television monitor or computer display that shows the entire scene continuously rather than just indicate the echoes from the region currently illuminated by the narrow antenna beam. Thus the role of the display has changed as the need for operator interpretation has decreased.

Types of Display Presentations The IEEE Standard on Radar Definitions includes 19 different types of display formats.[52] Most date to World War II and many are now seldom used. The standardized definitions do not cover all possible display formats. Given below are some of the more popular formats that have been employed. The IEEE uses the term "display" in its definitions but here we use either "scope" or "display" depending on what is perceived to be the more common usage. The following definitions are not precisely identical with the IEEE definitions, but they are consistent with them.

A-scope. *A deflection-modulated rectangular display in which the vertical deflection is proportional to the amplitude of the receiver output and the horizontal coordinate is proportional to range (or time delay).* This display is well suited to a staring or manually tracking radar, but it is not appropriate for a continually scanning surveillance radar since the ever-changing background scene makes it difficult to detect targets and interpret what the display is seeing.

B-scope. *An intensity-modulated rectangular display with azimuth angle indicated by one coordinate (usually horizontal) and range by the orthogonal coordinate (usually vertical).* It has been used in airborne military radar where the range and

angle to the target are more important than concern about distortion in the angle dimension.

C-scope. *A two-angle intensity-modulated rectangular display with azimuth angle indicated by the horizontal coordinate and elevation angle by the vertical coordinate.* One application is for airborne intercept radar since the display is similar to what a pilot might see when looking through the windshield. It is sometimes projected on the windshield as a heads-up display. The range coordinate is collapsed on this display so a collapsing loss might occur, depending how the radar information is processed.

E-scope. *An intensity-modulated rectangular display with range indicated by the horizontal coordinate and elevation angle by the vertical coordinate.* The E-scope provides a vertical profile of the radar coverage at a particular azimuth. It is of interest with 3D radars and in military airborne terrain-following radar systems in which the radar antenna is scanned in elevation to obtain vertical profiles of the terrain ahead of the aircraft. The E-scope is related to the RHI display.

PPI-display, or plan-position indicator. *An intensity-modulated circular display in which echo signals from reflecting objects are shown in plan view with range and azimuth angle displayed in polar (rho-theta) coordinates to form a map-like display.* Usually the center of the display is the location of the radar. A *sector-scan PPI* might be used with a forward-looking airborne radar to provide surveillance or ground mapping over a limited azimuth sector. An *offset PPI* is one where the origin (or location of the radar) is at a location other than the center of the display. This provides a larger display area for a selected portion of the coverage. The location of the radar with an offset PPI may be outside the face of the display.

RHI-display, or range-height indicator. *An intensity-modulated rectangular display with height (target altitude) as the vertical axis and range as the horizontal axis.* The scale of the height coordinate is usually expanded relative to the range coordinate. It has been used with meteorological radars to observe the vertical profile of weather echoes.

In addition, imaging radars such as synthetic aperture radar (SAR) and side-looking airborne radar (SLAR) generally display their output as a strip map with range as one coordinate and cross-range as the other coordinate. With the expanding graphics technology available from the computer industry, there is much more flexibility available in displaying radar information than previously.

Cathode Ray Tube Display The cathode ray tube (CRT), the origin of which dates back to the end of the nineteenth century, has been widely used as a radar display. There are two basic types of CRT displays. One is the *deflection-modulated CRT*, such as the A-scope, in which a target is indicated by the deflection of the electron beam. The A-scope displays the receiver output amplitude as a function of range, or time. An example was shown in Fig. 7.21, which illustrated the effect of frequency agility on clutter and target echoes. The other type is the *intensity-modulated CRT*, in which an echo is indicated by intensifying the electron beam and presenting a luminous spot on the face of

the CRT. An example is the PPI shown in Fig. 1.5. The TV CRT is also an example of an intensity-modulated display.

In general, a deflection-modulated display has the advantage of simpler circuits, and a target may be more readily discerned in the presence of noise, clutter, or interference. On the other hand, an intensity-modulated display, such as the PPI, has the advantage of presenting data in a more convenient and easily interpreted form. The deflection of the beam or the appearance of an intensity-modulated spot of a radar display caused by the presence of a target is commonly referred to as a *blip*.

Even though the CRT display has been widely used in radar, as well as in TV and in computers, it is by no means ideal. It employs a relatively large vacuum tube, which is a disadvantage compared to other types of displays. The entire display with its necessary circuits and controls can be even larger. The amount of information that can be presented is limited by the spot size of the electron beam. The number of resolution cells (pixels) per diameter might be one or two thousand, or more. In some high-resolution radars the number of resolvable range cells from the radar might be greater than the number of resolution cells available on the PPI. The result can be a collapsing loss. Increasing the CRT diameter does not necessarily increase the number of resolvable pixels since the spot diameter varies linearly with screen diameter. Another limitation of the intensity-modulated CRT is that its inherent dynamic range, or contrast ratio, might be of the order of 10 dB. This can cause blooming of the display by large targets so as to mask the blips from nearby smaller targets.

The decay characteristics of CRT phosphors are important when an operator views the screen to detect targets and extract information. The decay time of the visual information displayed should be long enough to allow the operator to detect the echo, yet short enough that the information painted on one scan does not interfere with the information entered from the succeeding scan. When processed information rather than raw video is displayed, the display characters might be dots, vectors to indicate direction and velocity, alphanumerics, or other appropriate symbols.

The improvements in electronic circuitry that have allowed major advances in signal and data processing have also benefited the CRT display. Character generators can be on a chip rather than occupy a bulky box. A complete deflection system can be placed on a chip. High-voltage power supplies that used to occupy a cubic foot and weigh 50 pounds[53] have been decreased considerably. Digital memories are small enough to replace the bulky analog scan converter. The required decay times for a PPI display need not depend on the decay characteristics of the phosphor, but an artificial persistence can be aehieved with electronic circuitry that controls the refresh rate of the display. In the past, CRTs often had to be viewed in a darkened room or with a hood, but the brightness of CRTs has been increased so that they can be used in ambient light or in sunlit aircraft cockpits. There have been significant advances in other types of displays due to the demands of TV and computers, but the CRT has been able to make significant advances as well. In spite of limitations, the CRT has been a competitive display because of its ruggedness, cost, color capability, ability to operate over wide temperature ranges, its wide viewing angle, and ability to conveniently display the type of information obtained by a radar.

In a conventional PPI display when raw, unprocessed video is displayed to an operator, some background noise should be present since it improves the operator's ability to

make a detection decision. A completely "black scope" has reduced sensitivity compared to one with some background noise. This applies to a radar with raw video and not to a display presenting processed data where the detection decision is made by automatic circuitry without the intervention of an operator.

Stroke and Raster Displays The conventional stroke PPI display is generated in synchronism with the rotating antenna rather than all at once as it might appear in a photo. The photo of a PPI display is usually made by opening the shutter of a camera and holding it open for one or more scans, as was done in Fig. 1.5. An operator viewing a normal PPI, on the other hand, sees a rotating radial line, or strobe, that rotates in synchronism with the scanning antenna. The trace of the rotating strobe with a raw-video display with no echoes present is normally dim, but is brightened to indicate the location of an echo signal when detected by the radar. The brightened blip fades with time depending on the persistence characteristics of the phosphor or the refresh characteristics of the electronic circuitry. The operator focuses his or her attention on the rotating radial strobe line to detect targets. This type of display is called a *stroke* display. In a stroke display the operator concentrates on that portion of the display in the vicinity of the strobe line since that is usually all that will be strong enough to be seen.

A TV-like display based on a raster scan* to provide a continuous picture of the radar output has some advantages over the stroke display. It can be made brighter than a stroke display. Information from other sensors such as other radars, the Air Traffic Control Radar Beacon System (ATCRBS), military identification friend or foe (IFF), low light level TV, forward-looking IR (FLIR), collision avoidance systems, or information from civil or military data links, can all be combined on one display. Maps of the region viewed by the radar as well as alphanumeric information and graphics can also be superimposed on a raster display, in addition to the processed radar video and raw video. A scan converter is required to change the format of the stroke display to that of a raster TV-like display.

Scan Converter[54] A scan converter changes the r,θ coordinates of a PPI into the x,y coordinates of a raster (TV-like) display. The r,θ coordinates are natural for the radar, but the x,y coordinates are more natural for viewing the output of the radar on a display. Early analog scan-converters were bulky, had low resolution, and poor presentation of gray scales. They were of limited utility compared to modern digitally generated scan converters in which the r,θ polar coordinates (range and azimuth) of the radar information are converted to rectangular x,y coordinates and stored in digital memory to generate a raster TV-like display. The raster display can be presented continuously to the operator since it can be refreshed at a rapid rate. If desired, an artificial decay can be inserted to imitate the decay characteristics of natural phosphors. Alternatively, there need be no decay and the image can be frozen for the time equal to the radar revisit time, and then updated. Displays with 2560 × 2048 pixels can be accommodated. The outputs of multiple radars can be shown together with appropriate symbology even though they may be air-surveillance radars with considerably different antenna rotation rates (revisit times) and weapon control radars with pie-shaped angular coverage of a limited sector. The use of a

*A raster is a scan pattern in which an area is scanned from side to side in lines from top to bottom.

scan converter usually does not seriously extend the latency of the display; that is, the time between echo detection and its display is held to a minimum. The format of a raster display can be that of a TV display or that of a computer monitor. The advantage of a TV-display format is that it can be recorded on tape with an inexpensive consumer video cassette recorder (VCR), viewed with a conventional TV monitor, and easily remoted using standard video cabling.

Flat Panel Displays (FPD)[55] Interest in the flat panel display for radar evolved from its successful development for commercial computer and TV applications. There have been several different types of FPDs produced or explored, but not all are suitable for radar.

The *liquid crystal display* (LCD) has been widely used for nonradar applications where low weight, volume, and power consumption are important as in laptop computers, watches, instruments, and calculators. The LCD does not emit light of its own, but operates by controlling the light that either passes through it or reflects off it. Usually the light is directed from behind, and the display is said to be *backlit*. There are two types of liquid crystal displays: the passive-matrix LCD and the active-matrix LCD. In the latter, a thin-film transistor is associated with each pixel of the display. The passive-matrix has seen much wider nonradar application than the active matrix because of its lower cost, but the active-matrix LCD (AMLCD) has much higher resolution, better image quality, it can display in color, and has faster response (greater video bandwidths). Thus the active-matrix LCD has more potential for radar applications than does the passive-matrix.

Other types of FPDs are the *plasma display* which can produce large flat full-color displays, electroluminescent displays, light-emitting diodes, and field emitter displays.

Flat panel displays such as the AMLCD and the plasma display have several important advantages over conventional CRTs. They are smaller, lighter, occupy less volume (reduced depth), and require less power than CRTs. In addition they are expected to have better reliability and reduced life-cycle cost. For most radar applications, however, they have to be more rugged than for commercial applications in that they usually have to withstand greater shock and vibration, as well as extremes in temperature.

The FPD is especially well suited for cockpit displays in military airborne applications and is replacing the CRT in many airborne systems.[56] In addition to presenting radar information, military cockpit displays must also handle data provided by electronic warfare sensors, command and control information for situation awareness, navigation information, alphanumeric data, graphics, and others.

Color in Radar Displays The availability of color in a radar display allows another "dimension" for the presentation of information. It can aid in providing a clear, easily understood picture of the situation as observed by radar. It is also an "attention getter" to alert the operator to something special or dangerous. Different colors can be used to indicate such things as the outputs of different radars presented on the same display; the outputs from multiple beams of a 3D radar; the areas of adverse weather with color coding of the weather by rain intensity; range rings; target tracks along with single detections; identification information from civil ATCRBS and/or military IFF; superimposed video maps of the area being observed by radar; and superimposed raw video. It can also be used to indicate the altitude or cross section of individual radar

Table 11.2 An Example of the Display Colors Used for an Airborne Weather Avoidance Radar[57]

Storm Intensity	Rainfall Rate (mm/h)	Rainfall Rate (dBZ)	Display Color
Drizzle	0.25 mm/h	13 dB	Black
Light	1.0 mm/h	23 dB	Green
Moderate	4.0 mm/h	33 dB	Yellow
Industry standard pilot alert	11.5 mm/h	40 dB	Red

The parameter dBZ was explained in Sec. 7.6. The "industry standard pilot alert" is the rainfall rate above which there might be hail that could damage an aircraft or cause sufficient turbulence to disturb passengers. Pilots are advised to stay clear of such areas.

echoes by color coding the target blips or by use of alphanumeric color symbols inserted on the display.

An example of the use of color is in the airborne weather-avoidance radar display, where the intensity of precipitation is designated by a distinctive color. A listing of the colors used by one radar manufacturer to indicate storm intensities is shown in Table 11.2.[57]

The original tricolor shadow-mask cathode-ray tube used for color TV did not have the resolution capability of monochrome displays or the penetration color tube which used a multilayer screen. This has changed with the increasing demands for high resolution computer color graphics as well as high-definition TV. Although a monochrome display with various shades of gray can be made to exhibit much of the same information that a color display can, color is capable of providing a greater number of distinguishable shades than can a monochrome display, is more pleasing, and has been widely accepted.

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PROBLEMS

11.1 (a) Find the overall noise figure of a superheterodyne receiver consisting of a low-noise RF amplifier with noise figure of 1.4 dB and gain of 15 dB, a mixer with 6.0-dB conversion loss and noise-temperature ratio of 1.2, and an IF amplifier with a noise figure of 1.0 dB. (b) What would be the noise figure of the receiver in (a) if the RF low-noise amplifier had a gain of 30 dB instead of 15 dB? (c) What would be the overall receiver noise figure if the IF amplifier noise figure in part (a) were 3.0 dB instead of 1.0 dB, and do you think this change is significant?

11.2 (a) Derive the overall noise figure of a receiver with noise figure F_r that is preceded by an RF device with a loss L_{RF}. (b) What is the overall noise figure of a transmission line and duplexer, which have a loss of 1.2 dB, connected to a receiver whose noise figure is 2.3 dB?

11.3 The greater the gain of the RF low-noise amplifier, the lower will be the overall noise figure. What adverse effect, however, occurs with an increase in the gain of the RF low-noise amplifier?

11.4 Show that the noise figure of a mixer is approximately the product of its conversion loss and the IF amplifier noise figure, when the diode mixer has a low noise-temperature ratio.

11.5 (a) Show that when a receiver of noise figure F_{RF} is attached to an antenna with antenna temperature T_a, the system noise figure F_s [Eq. 11.7] is

$$f_s = \frac{T_a}{T_0} + F_r$$

where T_0 is the standard temperature 290 K. (b) What is the system noise figure if the antenna temperature is 300 K, transmission line loss is 1.5 dB, and the receiver noise figure is 2.6 dB?

11.6 Consider a radar system with a receiver noise figure of 1.0 dB preceded by a transmission line with a loss of 0.5 dB. If the antenna temperature is 300 K, how important is it, in general, to attempt to reduce the receiver noise figure from 1.0 dB to 0.5 dB?

11.7 (a) Show that a radome with a loss L at a physical temperature T_{rd}, used with an antenna whose noise temperature in the absence of the radome is T_a', has an antenna noise temperature given by

$$T_a = \frac{T_a'}{L} + T_{rd}\frac{(L-1)}{L}$$

(b) Starting with the above, derive the *change* in antenna noise temperature $\Delta T_a = T_a - T_a'$ due to the presence of the radome. (c) Using the result of (a), what is the system noise figure? (d) What is the system noise figure when the receiver has a noise figure of 2.6 dB, a transmission line loss of 1.5 dB, an antenna with a noise temperature of 110 K, radome at a physical temperature of 310 K, and a loss of 0.6 dB through the radome?

11.8 (a) Find the noise bandwidth [Eq. (2.3)] of a network whose frequency response function $H(f) = (1 + jf/B)$, where B is the half-power bandwidth. (b) The above network is a single-stage low-pass RC filter. What is the expression for the noise bandwidth of a single-stage RLC bandpass network? (Do by inspection.) (c) Find the noise bandwidth for a low-pass filter with a gaussian shaped response $\exp[-a^2(f - f_0)^2]$, with $f > 0$.

11.9 Why might a double-conversion superheterodyne receiver be used instead of a single-conversion receiver? What limitation might there be in using a double-conversion receiver?

11.10 What effect does the local oscillator have on the receiver's dynamic range?

11.11 A receiver with a mixer front end has a noise figure of 6.6 dB. A low-noise amplifier (LNA) with a noise figure of 1.2 dB and gain of 10 dB is inserted ahead of the mixer to reduce the overall receiver noise figure. (a) How much of the new receiver noise figure is due to the mixer noise, and by how much has the dynamic range of the receiver been reduced? (b) If the gain of the LNA were increased to 20 dB, what would be the receiver noise figure and the decrease in dynamic range?

11.12 Why is a diode-limiter following the duplexer sometimes used as a receiver protector?

11.13 What duplexer options are available for a pulse doppler radar with a 10 percent duty cycle and a 0.1-μs pulse duration?

11.14 What limitations might there be in using an all-solid-state duplexer?

11.15 What is the usual cause, or criterion, for the end-of-life of a gas-discharge TR tube?

11.16 Consider a high prf pulse doppler radar with a 1-μs pulse width and a 10 percent duty cycle. (a) If a receiver protector is used that has a 1.5-μs recovery time, what fraction of the range coverage is blanked out? (b) If an electrostatic amplifier with a 30-ns recovery time is used, what fraction of the range coverage would be blanked out?